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METHOD AND APPARATUS FOR FRAME
ALIGNMENT IN DISCRETE MULTITONE TRANSCEIVERS

Field of Invention

The invention pertains to transceivers and modems. More

5   particularly, the invention pertains to asymmetric digital

subscriber line (ADSL) modems used to achieve very high speed

data communication via telephone networks.

Background of the Invention

There is an ever present desire to maximize the speed of

10   digital communications via networks, and particularly

telecommunication networks, such as public telephone systems.

Accordingly, telecommunication network providers now offer

customers many options for coupling to the telephone network in

addition to the standard analog based connection commonly

15   referred to as POTS (Plain Old Telephone System). Some of the

options that are widely available are Integrated Services Digital

Network (ISDN) lines, T-1 lines, E-1 lines, digital subscriber

lines (DSL) and asymmetric digital subscriber lines (ADSL).

ADSL's can provide very high data speeds, such as on the

order of several megabits per second, over a standard twisted

5   wire pair. Unlike the traditional data modems used for analog

communication with a telephone central office via a twisted wire

pair, ADSL requires modems both at the subscriber end and at the

telephone company Central Office end. Current ADSL systems

employ discrete multitone (DMT) technology to implement high

10   bandwidth communications, such as for digital TV broadcast, on

demand video, high speed video-based internet access, work at

home digital file transfer, teleconferencing, home shopping, and

other information services over existing twisted wire pair

telephone lines.

15   Several DMT standards have been promulgated. For instance,

the International Telecommunications Union (ITU) has promulgated

a standard for ADSL that is commonly termed G.lite and which is

set forth in ITU-T specification G.992.2, incorporated herein by

reference. Another standard, promulgated by ANSI, is commonly

20   termed Heavy ADSL and is set forth in ANSI specification T1.413,

issue 2, also incorporated herein by reference. In DMT

communications, data is sent in frames. A frame is comprised of

a plurality of samples, each frame including data samples and

cyclic prefix samples. Data samples comprise most of the frame

25   and the collection of data samples in a single frame comprise one

DMT symbol. The cyclic prefix is added at the beginning of each

frame and comprises the last L samples of that frame.

Accordingly, the cyclic prefix samples are between the DMT

symbols in the data stream. The purpose of the cyclic prefix is

5   to help avoid inter-symbol interference (ISI). The frame and

cyclic prefix is of a standardized length. For example, in heavy

ADSL, each symbol comprises 512 samples with 256 tones (32 tones

for upstream communications), each tone having a real and an

imaginary portion. Heavy ADSL utilizes a cyclic prefix of length

10   L=32 samples. Accordingly, a frame has 544 samples.

Figure 1 is a block diagram of the basic ADSL modem

functions, as would be well known to those of skill in the art.

The upper half of the diagram represents functions in the

transmit direction while the lower half represents functions in

15   the receive direction.

It should be noted that Figure 1 is a functional block

diagram and that the blocks shown therein do not necessarily

correspond to separate physical circuits. In fact, most if not

all of the functions will be performed by one or more digital

20   processors such as, but not limited to, a digital signal

processor, a micro processor, a programmed general purpose

computer, etc. It also is possible that part or all of the

functions of some of the blocks may be implemented by analog

circuitry.

In the transmit direction, digital data is transmitted from the transmitter 102 to a scrambler 104 that scrambles the data for transmission. The data is then processed through a forward error correction (FEC) encoder 106 which adds syndrome bytes to the data. The syndrome bytes will be used for error correction by the receiver at the receiving terminal. Next, as shown in block 108, the transmit data is encoded using quadrature amplitude modulation (QAM). The data is then converted from the frequency domain to the time domain via Inverse Fast Fourier Transform (IFFT) 110.

A 1:4 interpolator 112 interpolates the output of IFFT block 110 to produce 512 samples from the 128 samples output from block 110. The 32 sample cyclic prefix is added to each frame in block 114. The data is then forwarded to a coder/decoder (CODEC) 116. The CODEC encodes the data for transmission over the twisted wire pair to the receiving device.

In the receiver portion of transceiver 100, the received signal is passed from the twisted wire pair through the CODEC 116 where it is decoded. It is then passed to a time domain equalizer (TEQ) 118 to shorten the channel impulse response. Then, in 120, the cyclic prefix is removed. It will be appreciated by those of skill in the related arts that, if the length of the channel impulse response is less than or equal to the cyclic prefix length, then the Inter-Symbol Interference (ISI) can be eliminated by removing the Cyclic Prefix (CP)

length. Furthermore it is possible to compensate for channel

distortion with the frequency domain equalization (FEQ) in block

124, discussed further below.

   An echo canceller (EC) 134 is coupled between the transmit

5   path and the receive path and creates an echo cancellation signal

based on the transmit signal which is subtracted by subtractor

121 from the receive signal in order to cancel any echo of the

transmit signal that is present on the twisted wire pair that

could interfere with the received signal. The residual echo

10   signal is converted back to the frequency domain by fast fourier

transform (FFT) in block 122.

   The received signal, which now has had the cyclic prefix

removed and has been converted back to the frequency domain is

sent to block 124, where frequency domain equalization (FEQ) is

15   employed to compensate for the channel distortion .

   The received signal is then processed through a quadrature

amplitude modulation (QAM) decoder 126 to decode the tone signal

into digital data. That is followed by forward error correction

(FEC) in block 128 which uses the syndrome bits that were added

20   by the transmit path FEC encoder 106 to perform forward error

correction. Finally, the data is descrambled in block 130 to

extract the true data signal and then forwarded to a receiver

132.

   The output of the FFT block 122 also is sent to a timing

25   recovery circuit 136 that controls the CODEC 116 to synchronize

the CODEC to the timing of the received data. Essentially, the

timing recovery process is a feedback process in which the timing

tones are detected and used to continuously adjust the CODEC

timing so as to sample the received data at the appropriate

5   sampling points.

     In prior art frame alignment schemes, during

initialization, the transmitter transmits a known pattern in a

frame and the receiver attempts to receive that pattern. More

particularly, the transmitter transmits the same pattern in a

10   plurality of frames, for example 1024 consecutive frames. The

receiver performs a cross-correlation of the data on the two wire

pair with the expected pattern while the transmitter transmits

those 1024 frames and determines which set of L+1 consecutive

samples yields the peak cross correlation. The timing that

15   resulted in the peak cross correlation calculation is selected as

the start time of a frame or a symbol. It means that the sample

that resulted in the peak cross correlation calculation is the

first sample of the frame, e.g., 544 samples (before removing the

cyclic prefix).

20   This technique for frame alignment or symbol alignment is

computationally demanding and is not as accurate as desired.

     Accordingly, it is an object of the present invention to

provide an improved method and apparatus for frame alignment in a

DMT transceiver.

It is another object of the present invention to provide an improved DMT transceiver.

Summary of the Invention

The invention is a method and apparatus for aligning with a

5   transmit frame in a discrete multitone (DMT) transceiver so as to minimize inter-symbol interference (ISI). The method and apparatus involves calculating the shortening channel impulse response and then determining the particular set of consecutive samples of length L +1 in the shortening channel impulse

10 response, that has a maximal energy. Once this particular set of coefficient is determined, the first coefficient of that set is selected as a starting location h'_0 in shortening impulse response for determining frame alignment. This can be achieved by skipping samples of the receive signal in a particular frame.

15 The majority of the interference of the immediately preceding symbol on the present symbol typically is approximately contributed by shortening channel impulse response coefficients h'_{L+1}, h'_{L+2}, ... h'_{P-1}, where P is the DMT symbol length (e.g., 512 samples). This is true because the channel impulse response

20 coefficients, h'_0, h'_1, ... h'_L, do not contribute to the ISI. Therefore, the majority of the interference can be calculated with the shortening channel impulse response coefficients h'_{L+1}, h'_{L+2}, --- and h'_{P-1}.

The noise on the i-th subcarrier is a complex value r_i+s_ij,

where $j=\sqrt{-1}$, and the absolute value $\sqrt{r_i^2 + s_i^2}$ is considered to be

the ISI noise on the i-th subcarrier. The total ISI noise is

computed by summing all the inner products of the weight factor

vector W and the ISI leakage energy vector from the previous

symbol, $\overline{C}|h_i'|FV_k$. The rotation of the shortening channel impulse

response, that yields the smallest total ISI noise will be chosen

as the frame alignment point.

Accordingly, to determine the best rotation of the

shortening channel impulse response coefficients for frame

alignment, we calculate the ISI using the rotation as described

above as well as several immediately surrounding rotations of the

coefficients and select the rotation that yields the lowest ISI

value. The calculations are performed in the frequency domain.

Specifically, we multiply the average energy of the transmitted

symbol (which is a constant) by the Fast Fourier Transform (FFT)

of the ISI value caused by the imperfect shortening channel

impulse response coefficients. We do this for various rotations

including and immediately surrounding the rotation determined

above and take the rotation that yields the smallest value. It

has been found that excellent results can be achieved by

calculating the interference value for a relatively small number

of rotations. For instance, in a DMT system with 512 samples per

symbol and a 32 sample cyclic prefix, excellent results can be

achieved by calculating interference for about seven different

rotations surrounding and including the rotation corresponding to

5   the maximum channel response.

Brief Description of the Drawings

        Figure 1 is a block diagram illustrating the functions of an

ADSL modem transceiver in accordance with the prior art.

        Figure 2 is a block diagram illustrating in more detail

10   channel shortening in discrete multitone communications in

accordance with the prior art.

        Figure 3 is a block diagram illustrating the functions of an

ADSL modem transceiver for digital multitone communications in

accordance with the present invention.

15   Figure 4 is a flowchart illustrating symbol alignment in

accordance with the present invention.

Detailed Description of the Invention

        Figure 2 is a block diagram that helps illustrate time

domain equalization in DMT in more detail. In particular, block

20   203 encompasses functions performed at the transmitter. Block

207 encompasses functions that occur in the data channel and

block 213 encompasses functions performed at the receiver. Block

205 encompasses those functions which result in channel

shortening. Thus, it can be seen that, in the transmitter 203, a DMT signal, x(n), is generated for transmission via the channel. The channel 207 has a channel impulse response, h(n), represented by block 209. Accordingly, the signal, y(n), received at the

5   receiver 213 is the convolution of the original transmitted signal x(n) and the channel response h(n), i.e.;

$$y(n) = x(n) * h(n) \qquad\qquad \text{Eq. (1)}$$

The time domain equalization circuit (TEQ) 211 in the receiver 213 is a short time-domain finite impulse response (FIR)

10   filter that convolves the received signal, y(n), with a coefficient 1+a(n). The output y'(n) from the time domain equalization domain circuit 211 therefore is :

$$y'(n) = y(n)*(1+a(n)) = x(n)*(h(n)*(1+a(n)))=x(n)*h'(n) \qquad \text{Eq. (2)}$$

Let us call h(n) * (1+a(n)) the shortening channel impulse

15   response (or shortening channel) and denote it as h'(n) having a length M.

The TEQ circuit 211 (i.e., Eq. (2)) shortens the channel impulse response to the length of the cyclic prefix, e.g., 32 samples. Then the cyclic prefix is removed from the signal y'(n)

20   in block 215.

If the shortening channel length, M, is less than or equal to the length, L, of the cyclic prefix, i.e., M≤L, then each y'(n) will have no dependence on any transmitted samples before x(n-L), i.e., there will be no inter-symbol interference (ISI).

Thus, after the cyclic prefix is removed, the actual symbol will remain without inter symbol interference. However, if M>L, then the sampling signal y'(n) is dependent on the samples before x(n-L) and, thus, the removal of the cyclic prefix would not prevent inter-symbol interference and thus would degrade the performance of the transceiver. Hence, it is desirable to design an efficient TEQ shortening filter, 1+a(n), that makes M≤L and for which the energy in the channel impulse response h'(n) with 0 ≤ n < 512 is as small as possible for n≥L and as large as possible for n<L.

For proper operation of the transceiver in accordance with the description above, of course, the frame in the receiver must align itself to the frame in the remote transmitter in order to read the frame samples correctly. The proper frame alignment point is determined during initialization.

U.S. Patent Application No. 09/481,027 entitled "Method and Apparatus for Time-Domain Equalization in Discrete Multitone Transceivers", filed on January 11, 2000, describes a TEQ shortening filter that produces an overall channel response h'(n) for which M≤L, as desired in order to eliminate ISI.

The impulse response h(n) of the channel has the transfer function:

$$H(z) = \frac{B(z)}{A(z)} = \frac{b_0 + b_1 z^{-1} + b_2 z^{-2} + \ldots + b_{(L-1)} z^{-(L-1)}}{1 + a_1 z^{-1} + a_2 z^{-2} + \ldots + a_{(K-1)} z^{-(K-1)}} \qquad \text{Eq. (3)}$$

where L is the length of the cyclic prefix, and K is the length
of the TEQ filter having impulse response (1+a(n)) or A(z). In
accordance with equation (3), the transfer function of the
shortening channel is:

$$\bar{h}(z) = A(z)H(z) = b_0 + b_1 z^{-1} + b_2 z^{-2} + \ldots + b_{(L-1)} z^{-(L-1)} \quad \text{Eq. (4)}$$

Thus, the length, M, of the shortening channel is set equal to
the length of the cyclic prefix, L, so that ISI between two
consecutive symbols can be eliminated to improve the overall
performance of the transceiver by removing the cyclic prefix.

Let

$$I = [b_0 \ b_1 \ b_2 \ \ldots \ b_{(L-1)} \ -a_1 \ -a_2 \ldots -a_{(K-1)}]^T = [b \quad a]^T \quad \text{Eq. (5)}$$

and

$$W(n) = [x(n) \ x(n-1) \ x(n-2) \ldots x(n-(L-1)) \ y(n-1) \ y(n-2) \ldots y(n-(K-1))]^T \quad \text{Eq. (6)}$$

where T denotes the transpose of a matrix. From equations (3)
through (6), the estimated signal, $\bar{y}(n)$, at the receiver is:

$$\bar{y}(n) = I^T W(n) \quad\quad\quad\quad\quad \text{Eq. (7)}$$

and the error is:

$$e(n) = y(n) - \bar{y}(n) \quad\quad\quad\quad\quad \text{Eq. (8)}$$

From the well known LMS (Least Mean Square) error algorithm, the
optimal solution I_{opt} is

$$RI_{opt} = r \quad\quad\quad\quad\quad \text{Eq. (9)}$$

where

$$R = E\{W(n)W^T(n)\} \qquad\qquad \text{Eq. (10)}$$

$$r = E\{y(n)W(n)\} \qquad\qquad \text{Eq. (11)}$$

$E\{\}$ in the equations above represents a generalized averaging

5   function or an ensemble average.

Since some channel interference is caused by channel noise and

impairments, it is very difficult to achieve a shortening channel

impulse response that has only (L+1) continuous nonzero values

and zeros in the remaining slots. Hence, accurate frame

10   alignment can be very critical to the overall performance of DMT

transceivers.

Let $[h_0', h_1', h_2', ..., h_{511}']$ be the shortening channel impulse

response. As previously mentioned, one way to locate the frame

alignment roughly is by finding the series of consecutive

15   coefficients of length L+1 (L is the channel prefix length) that

yields the maximal energy. By skipping samples in the receive

signal in a particular frame, we can allocate the first

coefficient of the series of consecutive coefficients of length

L+1 as h'_0, whereby the coefficients from h'_0 to h'_L yield the

20   maximal energy. The coefficients of the shortening channel

impulse response from h'_{L+1} to h'_{P-1} contribute to 1S1. Hence, we

try to make the energy as low as possible for h'_{L+1} to h'_{P-1}.

Although there is maximal energy in h'_0 to h'_L, it is not

necessarily the best choice for frame alignment. Rather, we know

25   only that the frame alignment point is close to this point.

Thus, we implement a fine-tuning algorithm in accordance with the present invention to find the best location for frame alignment.

Let $c = [c_0, c_1, c_2, \ldots c_{511}]$ and $d = [d_0, d_1, d_2, \ldots d_{511}]$ be the samples of previous symbol and current symbol, respectively, that are sent from the transmitter. Due to the addition of the 32 samples of cyclic prefix, the samples that go out through the CODEC and channel will be $[c_0, c_1, c_2, \ldots c_{511}, d_{480}, d_{481}, \ldots d_{511}, d_0, d_1, d_2, \ldots d_{511}]$. Let the corresponding samples at the receiver be $[c'_0, c'_1, c'_2, \ldots c'_{511}, d''_{480}, d''_{481}, \ldots d''_{511}, d'_0, d'_1, d'_2, \ldots d'_{511}]$. Thus, after removing the cyclic prefix, we have:

$$d'_0 = d_0 h'_0 + d_{511} h'_1 + d_{510} h'_2 + \ldots + d_{480} h'_{32} + c_{511} h'_{33} + \ldots + c_{33} h'_{511} \qquad (12)$$

$$d'_1 = d_1 h'_0 + d_0 h'_1 + d_{511} h'_2 + \ldots + d_{480} h'_{33} + c_{511} h'_{34} + \ldots + c_{32} h'_{511} \qquad (13)$$

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$$d'_{511} = d_{511} h'_0 + d_{510} h'_1 + d_{509} h'_2 + \ldots + d_0 h'_{511} \qquad (14)$$

From equations (12) to (14) and the imperfectly shortening channel impulse response, the previous symbol will interfere with the current symbol. The interference is $c_{511} h'_{33} + \ldots + c_{33} h'_{511}$ in d'_0, $c_{511} h'_{34} + \ldots + c_{32} h'_{511}$ in d'_1, etc. If we assume that the transmitted signal is a pseudo random and stationary signal, the major interference from the previous symbol on the current symbol is $h'_{33}, \ldots, h'_{511}$. That is:

$$INTF = \overline{C}[|h'_{33}|+|h'_{34}|+|h'_{35}|+...+|h'_{511}|,|h'_{34}|+|h'_{35}|+|h'_{36}|+...+|h'_{511}|,..............,|h'_{511}|,\underbrace{0,...,0}_{33}] \quad (15)$$

where $\overline{C}$ is the average energy value of the transmitted symbol

and INTF represents the symbol interference.   Equation (15) can

be rewritten as:

$$INTF = \overline{C}(|h'_{33}|[\overbrace{1,0,0,...,0}^{512}] + |h'_{34}|[\overbrace{1,1,0,....,0}^{512}] + |h'_{35}|[1,1,1,0,...,0]+.....$$

$$|h'_{511}|[\overbrace{\underbrace{1,1,1,...,1}_{479},\underbrace{0,...,0}_{33}}^{512}]) \quad (16)$$

In equation (16), $\overline{C}$ is a constant and can be ignored.   The

coefficient of the h'₅₁₁ shortening channel impulse response

contributes more symbol interference that the coefficient h'_{33}

and the coefficients from h'_0 to h'_{32} do not affect the symbol

interference at all due to the removal of the cyclic prefix.

   In accordance with the analysis above, the symbol

interference of the previous symbol on the current symbol is

caused by the nonzero coefficients in the shortening channel

impulse response.   In order to reduce the symbol interference, we

rotate the shortening channel impulse response $[h'_0, h'_1, h'_2, ..., h'_{511}]$

such that the last 479 (512-33) coefficients of the shortening

channel impulse response have less symbol interference on the

next symbol.

5   Let us define V_k as

$$V_k = [\underbrace{1,1,...,1}_{k1s}, \underbrace{0,...,0}_{(512-k)0s}].$$

Then equation (15) and (16) can be rewritten as

$$INTF = [intf_0, intf_1, ..., intf_{511}]$$
$$= \bar{C}(|h'_{33}|V_1 + |h'_{34}|V_2 + ... + |h'_{511}|V_{479}) \tag{17}$$

We can determine how much effect the symbol interference has

10   on each tone in the receiver by taking the FFT (Fast Fourier

Transform) of equation (16), as follows;

$$FINTF = FFT_{512}([intf_0, intf_1, ..., intf_{511}])$$
$$= \bar{C}(|h'_{33}|FV_1 + |h'_{34}|FV_2 + ... + |h'_{511}|FV_{479}) \tag{18}$$

where

$$FFT_{512}(V_k) = (r_o + s_o j, r_1 + s_1 j, ..., r_{255} + s_{255} j) \text{ and}$$

$$FV_k = [t_o, t_1, ..., t_{255}] \text{ is a 256 real value vector with } t_i = \sqrt{r_i^2 + s_i^2}.$$

In ADSL, the different tones will be loaded by different numbers

of bits, depending on the tone's SNR (Signal to Noise Ratio).

The noise is caused by ISI and all other effects of line

impairments (such as echo, noise from radio signals, etc.) A bit

loading algorithm will assign a certain number of bits to each

tone based on the SNR, the capacity of bit rate, and other

factors. For more generality, in equation (18), the noise on

each tone can be weighted by a different value, W_0, W_1, ..., W_{255}.

In that case, the weight factor vector W on the 256 tones would

be the vector.

$$W = [W_0, W_1, ..., W_{255}]^T \tag{19}$$

Then the total noise, TN, caused by ISI can be expressed as

$$TN = | \text{ total ISI noise } | = | \text{ FINTF·W } |$$

$$= \bar{C}(|h'_{33}|(FV_1 \cdot W) + |h'_{34}|(FV_2 \cdot W) + ... + |h'_{511}|(FV_{479} \cdot W)) \tag{20}$$

where

$|x|$ denotes the absolute value of x,

FV_i is a vector of 256 real values, and

$FV_i \cdot W$ is an inner product of FV_i and W.

The weight factors W_i in equation (19) could be different for

each of the 256 tones, depending on the importance of the tone

based on its location. For example, the weight factors for the

tones from 0 to 31 can be set to zeros since they are used for

transmitted signals. Also, the weight factor for the timing tone

(e.g., W_{63}) can be set slightly higher than the weight factor of

the other tones since tone 63 is an important tone because it is

involved in the performance of timing recovery. If timing

recovery performance is poor, it will degrade the overall

performance of the system and affect all of the tones.

Equation (20) exactly describes the symbol interference by

the nonzero coefficients $h'_{33},\ldots,h'_{511}$ in the shortening channel

impulse response on each tone in the receiver. Since the first

33 samples, from h_0' to h_{32}', in the shortening channel impulse

response do not cause symbol interference, we can start to find

the maximal energy of L+1 (33) consecutive coefficients of the

shortening channel impulse response by rotating the shortening

channel impulse response coefficients in a particular frame such

that the coefficients from h_0' to h_{32}', have the maximal energy.

After that, we employ equation (20) to fine-tune to the best

rotation around this range, i.e., the rotation that yields the

minimal TN value.

For instance, if we want to consider this rotation plus the

three rotations to each side of this rotation (for a total of

seven tested rotations), then we would perform equation (20) a

total of seven times, as follows:

$$TN_1 = |FINTF_1 \cdot W| = \bar{C}(|h'_{30}|(FV_1 \cdot W) + |h'_{31}|(FV_2 \cdot W) + |h'_{32}|(FV_3 \cdot W) + \dots$$
$$+ |h'_{507}|(FV_{478} \cdot W) + |h'_{508}|(FV_{479} \cdot W))$$

$$TN_2 = |FINTF_2 \cdot W| = \bar{C}(|h'_{31}|(FV_1 \cdot W) + |h'_{32}|(FV_2 \cdot W) + |h'_{33}|(FV_3 \cdot W) + \dots$$
$$+ |h'_{508}|(FV_{478} \cdot W) + |h'_{509}|(FV_{479} \cdot W))$$

$$TN_3 = |FINTF_3 \cdot W| = \bar{C}(|h'_{32}|(FV_1 \cdot W) + |h'_{33}|(FV_2 \cdot W) + |h'_{34}|(FV_3 \cdot W) + \dots$$
$$+ |h'_{509}|(FV_{478} \cdot W) + |h'_{510}|(FV_{479} \cdot W))$$

$$TN_4 = |FINTF_4 \cdot W| = \bar{C}(|h'_{33}|(FV_1 \cdot W) + |h'_{34}|(FV_2 \cdot W) + |h'_{35}|(FV_3 \cdot W) + \dots$$
$$+ |h'_{510}|(FV_{478} \cdot W) + |h'_{511}|(FV_{479} \cdot W))$$

$$TN_5 = |FINTF_5 \cdot W| = \bar{C}(|h'_{34}|(FV_1 \cdot W) + |h'_{35}|(FV_2 \cdot W) + |h'_{36}|(FV_3 \cdot W) + \dots$$
$$+ |h'_{511}|(FV_{478} \cdot W) + |h'_0|(FV_{479} \cdot W))$$

$$TN_6 = |FINTF_6 \cdot W| = \bar{C}(|h'_{35}|(FV_1 \cdot W) + |h'_{36}|(FV_2 \cdot W) + |h'_{37}|(FV_3 \cdot W) + \dots$$
$$+ |h'_0|(FV_{478} \cdot W) + |h'_1|(FV_{479} \cdot W))$$

$$TN_7 = |FINTF_7 \cdot W| = \bar{C}(|h'_{36}|(FV_1 \cdot W) + |h'_{37}|(FV_2 \cdot W) + |h'_{38}|(FV_3 \cdot W) + \dots$$
$$+ |h'_0|(FV_{478} \cdot W) + |h'_1|(FV_{479} \cdot W))$$

The one of these seven values for TN that has the smallest value

5   is selected as the best rotation to use for frame alignment.

For example, if TN_4 is the minimal ISI, then there is no need to do anything further because the original frame alignment is the best one. However, if TN_3 is the minimal ISI, then the receiver should skip 511 (i.e., 512-1) samples in one frame in order to

5   rotate the shortening channel impulse response as follows:

$$h_0' \rightarrow h_1', h_1' \rightarrow h_2', ..., h_{511}' \rightarrow h_0'.$$

In practical implementations, the $FV_i \cdot W$ computations can be stored in tables to reduce processing time.

Figure 3 is a block diagram of a DMT transceiver in

10   accordance with the present invention. Since the present invention pertains to the determination of proper frame alignment during initialization, all of the blocks shown in prior art Figure 1 remain the same and are shown in Figure 3 bearing the same reference numerals. The invention resides primarily in

15   frame alignment blocks 140 and 146 and switches 142 and 144. Particularly, the processing described above is performed during initialization on the received data. Switches 142 and 144 are set to cause the received data to bypass blocks 140 and 146 during normal operations.

20   In operation, block 140 first cross-correlates the received data with a pre-defined periodic transmitted pattern frame signal to find the real channel impulse response without shortening the channel. This process results in a coarse location of the frame alignment position and prepares for TEQ training. TEQ training

25   is then performed. After the completion of TEQ training, the

shortening channel impulse response is determined in block 146 in

accordance with equation 20 to fine-tune the correct sampling

position for final frame alignment.

The process performed in block 146 is illustrated in the

flow chart of Figure 4. The process starts at step 400. In step

402, the shortening channel impulse response h', is determined in

accordance with well known mathematical steps. Then, in step

404, the shortened impulse response is queried to find the set of

L+1 consecutive samples that has the maximum energy. Variable i

is then set to the number of the first sample of that set of

samples (i.e., the shortened channel impulse response is rotated

to the station for which the first).

In step 406, counter n is set to 1 and i is set to i-3.

Note that the value of i must be calculated modulo P, where P is

the shortening channel impulse response length, e.g., 512 for

heavy ADSL. For example, if, in step 404, i is set to 2, then i-

3 is 511 when counting modulo 512. In step 408, equation (20) is

used to calculate TN for this particular rotation of the samples.

The result is saved as TN_n. In the example that is illustrated

in Figure 4, TN will be calculated seven times, namely, for those

locations starting at i-3, i-2, i-1, i, i+1, i+2, and i+3.

Accordingly, in step 412, it is determined whether TN has been

calculated for all seven rotations. If not, flow proceeds to

step 410 where both i and n are incremented and flow returns to

step 408 where TN is calculated for the next rotation.

When it is determined in step 412 that TN has been
calculated for all seven rotations, flow proceeds to step 414.
In step 414, the rotation that yielded a minimum value for TN
(for the seven rotations tested) is selected as the rotation to

5   be used for frame alignment during normal operation. The process
ends at step 416.

It will be understood by those of skill in the art that n
can be set to any other reasonable value. It has been found that
the best rotation for frame alignment (as determined by equation

10   20) typically will be within about three samples of the rotation
corresponding to the maximal channel impulse response energy in
L+1 consecutive coefficients as determined in step 404.
Accordingly, n=7 is a reasonable choice. However, it can be any
number less than 512. For general application, the weighed

15   vector W in equation 19 can be set to

$$[\underbrace{0,...,0}_{32},\underbrace{1,...,1}_{480}].$$

However, it can be set to provide a different value for each tone
if desired.

Having thus described a few particular embodiments of the

20   invention, various alterations, modifications, and improvements
will readily occur to those skilled in the art. Such
alterations, modifications and improvements as are made obvious
by this disclosure are intended to be part of this description
though not expressly stated herein, and are intended to be within

the spirit and scope of the invention. Accordingly, the

foregoing description is by way of example only, and not

limiting. The invention is limited only as defined in the

following claims and equivalents thereto.